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HF ANTENNA SWITCH

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CONTENTS

1. INTRODUCTION.....3

2. OPERATING INSTRUCTIONS.....4

3. TESTING PROCEDURES.....5

APPENDICES

A. A Switched Radar Calibration Antenna.....11

B. Fibre Optics.....19

C. Antenna Switch Design.....21

D. Estimation of Capacitance and Inductance.....23

E. Circuit Diagram.....25

F. Printed Circuit Layout.....26

G. Mechanical Drawings.....27

H. Summary of Data Sheets.....32

1. INTRODUCTION

This switching unit has been developed as a principal component of a radar target comprising a half wave dipole antenna with a switch at its centre. The switch causes a modulation of the Radar Cross-Section (RCS) which can be interpreted by the radar as a target Doppler shift. For completeness, the theory is provided in Appendices A to D, which are proprietary to London Research and Development Corporation and are not a part of the contract.

The unit embodies a semiconductor MOSFET switch which is driven by a square wave oscillator based on the "555" timer chip. It is powered by two 9V transistor radio batteries (e.g. Duracell type MN1604-6LR61) in parallel which have a total capacity of about 0.12 amp-hr. The circuit draws about 10 mA so that it can switched on and left unattended for several hours before the batteries require changing. The MC1455 timer chip is very insensitive both to the line voltage and to temperature.

The housing is constructed from solid blocks of Delrin plastic which have been machined to size. Delrin is easy to machine and remains very tough over a wide range of temperatures specifically from -30 C to 85 C. Its dielectric constant is about 3.7. The housing consists of a container for the printed circuit board and batteries, a neoprene seal, a lid and a mounting plate. The antenna terminals are threaded into the container and, if required may be sealed with gasket cement. Thus there should be no access for water or dust and the assembled unit is weatherproof.

The switching electronics is designed to function over a wide range of radio frequencies from less than 1 MHz to over 25 MHz. The critical element is the semiconductor switch, which must have a low impedance when in the "ON" position and a very high impedance when it is "OFF". (These impedances must be compared to the impedance of the antenna itself which should be principally resistive and be practically equal to the radiation resistance at resonance, namely about 75 Ohm.) A MOSFET has been chosen for the switch and this is characterized by its resistance when in the conducting state and its capacitance when it is not conducting. The dc resistance between the antenna terminals when "ON" is less than 2 Ohm while capacitance across the terminals of the unit when "OFF" is about 50 pF. At high frequencies, the impedance of the switch in the non-conducting state is increased by placing an inductor across it to cancel the capacitive reactance.

There are three banks of switches inside the unit and three jumper positions. The first switch is the main power switch for the device. This is a single unit with labels "On" and "Off". The 8-position DIP switch controls the oscillator frequency while the 4-position DIP switch sets the value of the inductance in parallel with the semiconductor antenna switch.

The position of the jumper shunt determines how the semiconductor antenna switch is driven. When the shunt is in the central position on JP-2, the MOSFET gate is connected to the oscillator. When it is across JP-1, the gate is connected to the positive rail and the antenna switch conducts. When it is across JP-3, the gate is connected to "ground" (the negative battery

terminals and the FET source). In this position the FET is non-conducting.

Some protection for the MOSFET is provided in the form of a neon wired directly from source to drain. This reacts to spikes of nanosecond duration by striking at about 85 V. For some types of MOSFET this will prevent surges exceeding the maximum source-drain voltage and may assist even when the transistor specifications are exceeded by spikes of short duration.

There is also a fuze which will blow if large currents flow through the neon (or FET). The resistance of small current fuzes can badly affect the performance of the antenna switch so that the fuze supplied is actually a 1 A fuze and, in this case, it will probably be ineffective for protection against high currents.

If protection is a serious issue, the MOSFET, which is typically a 2N7000, must be changed to a Motorola BS107. Data sheets are summarised in Appendix H.

2. OPERATING INSTRUCTIONS

When the unit must operate close to strong sources of interference, the standard switching transistor, which is a 2N7000, must be changed to a BS107. This requires removal of the circuit board and the replacement of the transistor.

If it should be necessary to dismantle and reassemble the unit, care must be exercised with the screws by *avoiding too much torque*. This can strip the threads in the Delrin or mar the surface. Similarly, *do not apply excessive torque* to the antenna connection screws as this may loosen the terminals in the plastic.

The Delrin container can be opened by removing the six wing nuts holding the lid in position. The lid is apt to stick to the neoprene gasket and it can be loosened by *carefully compressing* the gasket material at either end by inserting a screwdriver blade: a slight rotation of the blade should separate the lid from the container without marking the plastic.

The state of the batteries can be determined by placing the probes of a voltmeter on the battery probe areas on the green circuit board. These probe points are labelled "BAT+" and "BAT-". When the unit's On-Off switch is in the Off position, the open circuit battery voltage still appears across the probe points. When it is placed in the On position, the voltage can be measured under load. If this voltage is below about 8.5 V, only an hour or so of operation is possible and the batteries should probably be changed.

The jumper shunt should be set across JP-2, which is the central jumper position. The operation of the oscillator can be monitored by attaching an oscilloscope probe between the "FREQ" probe point and "BAT-". A square wave going from close to 0 V to approximately 7 V should appear on the screen. The frequency of this square wave is set from the 8-position DIP

switch (typically coloured black). When all the switches in this bank are set to Off, the oscillator frequency is nominally 1 kHz, which is well into the audio range. This setting was included to facilitate an aural test of the system using radio stations but may be of limited use.

The 4-position DIP switch is set according to the radio-frequency When the radio-frequency is less than 10 MHz, all the switches are set in the Off position. For frequencies between 10 MHz and 17 MHz switch 4 should be On and the remainder Off. For 18 MHz to 22 MHz set only switch 3 On. From 22 MHz to 24 MHz, only position 2 should be on. For frequencies above 24 MHz, set only position 1 to the On position.

In the event that the antenna is not cut exactly to one half wavelength overall, the inductors can be employed to tune the antenna impedance in which case it is possible to depart from the previous settings.

The settings are summarised in the following tables.

TABLE 1
JUMPER SETTINGS

Jumper	Purpose
JP-1	Antenna Switch set to closed
JP-2	Antenna Switch connected to Oscillator
JP-3	Antenna Switch set to open

TABLE 2 OSCILLATOR FREQUENCIES 8-POSITION DIP SWITCH	
SWITCH STATE	NOMINAL FREQUENCY (Hz)
ALL OFF	1000
SW-1 ON, Remainder OFF	0.25
SW-2 ON, Remainder OFF	0.5
SW-3 ON, Remainder OFF	1.0
SW-4 ON, Remainder OFF	2.0
SW-5 ON, Remainder OFF	4.0

TABLE 3 INDUCTOR SWITCH SETTINGS	
SWITCH STATE	EFFECT
ALL OFF	No inductors connected. Operation at frequencies less than 10 MHz
SW-1 ON, Remainder OFF	0.68 µH connected. Operations about 25 MHz.
SW-2 ON, Remainder OFF	1.00 µH connected. Operations about 22 MHz.
SW-3 ON, Remainder OFF	1.2µH connected. Operations about 18 MHz.
SW-4 ON, Remainder OFF	2.2 µH connected. Operations from 10 to 15 MHz.

To seal the container, the gasket and lid can be replaced and the wing nuts re-attached. The battery wires *should not be trapped* under the lid. The unit should be rendered weatherproof by tightening the wing nuts moderately by hand: *do not use excessive torque* as this can compress the gasket unnecessarily or distort the lid.

It is important that the *batteries are not allowed to run-down completely* because this may cause them to swell and distort or damage the container.

In the event that the oscillator frequency requires trimming, there are places in the printed circuit board for additional capacitances.

3. TESTING PROCEDURES

The unit (with the 2N7000 transistor) has been tested by applying a radio frequency signal to the terminals of the unit through a resistor as shown in Figure 1. The RF voltage across the terminals has been compared to that from the signal generator using an oscilloscope.

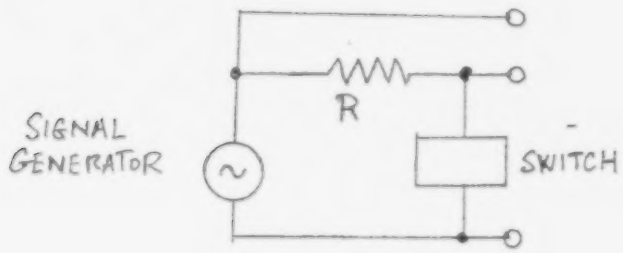

Figure 1.
Test Circuit

With the antenna switch configured in the open position either by connecting the jumper shunt across JP-3 or by keeping the oscillator switched Off, the impedance of the switch is high. The measurements in this condition were made with a series resistor, R, of 470 Ohm. Figure 2 shows a graph of the voltage ratio as a function of frequency. This suggests that, over most of the frequency range, the impedance is greater than 470 Ohm and is usually significantly greater.

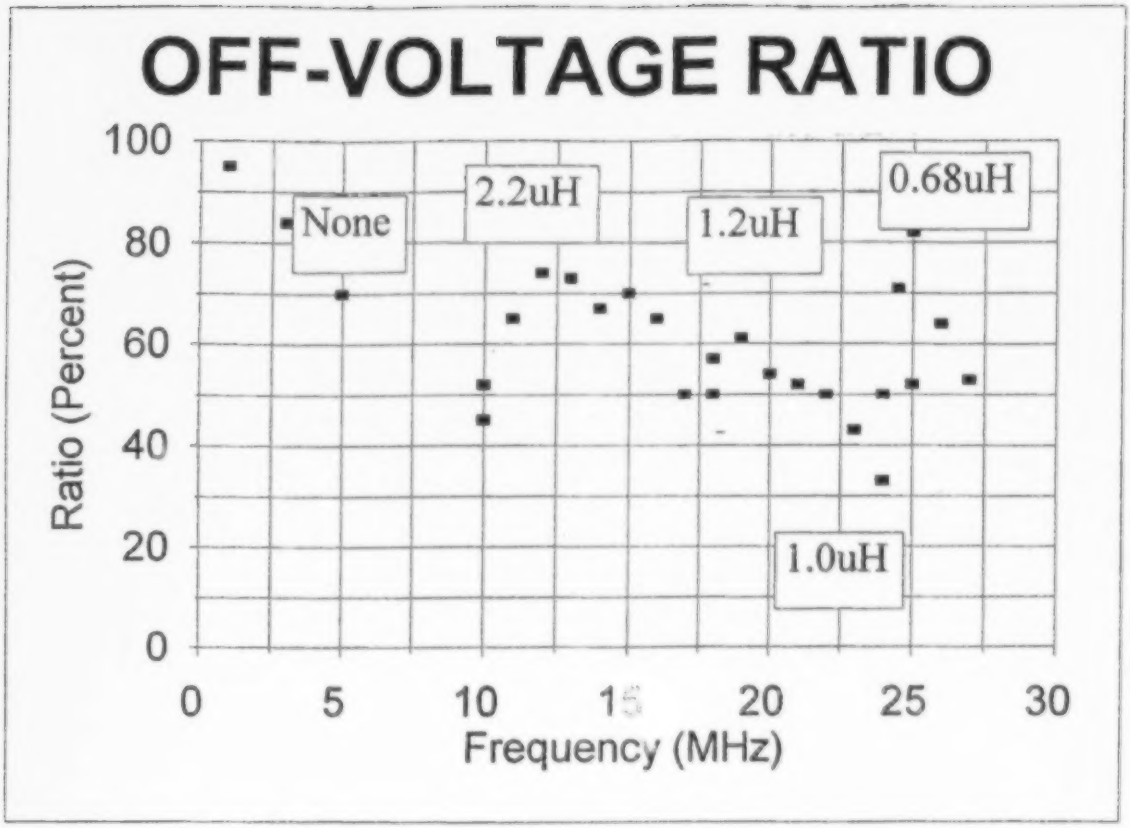
OFF-VOLTAGE RATIO
Ratio (Percent)
100
80
60
40
20
0
None
2.2uH
1.2uH
0.68uH
1.0uH
0
5
10
15
20
25
30
Frequency (MHz)

Figure 2.

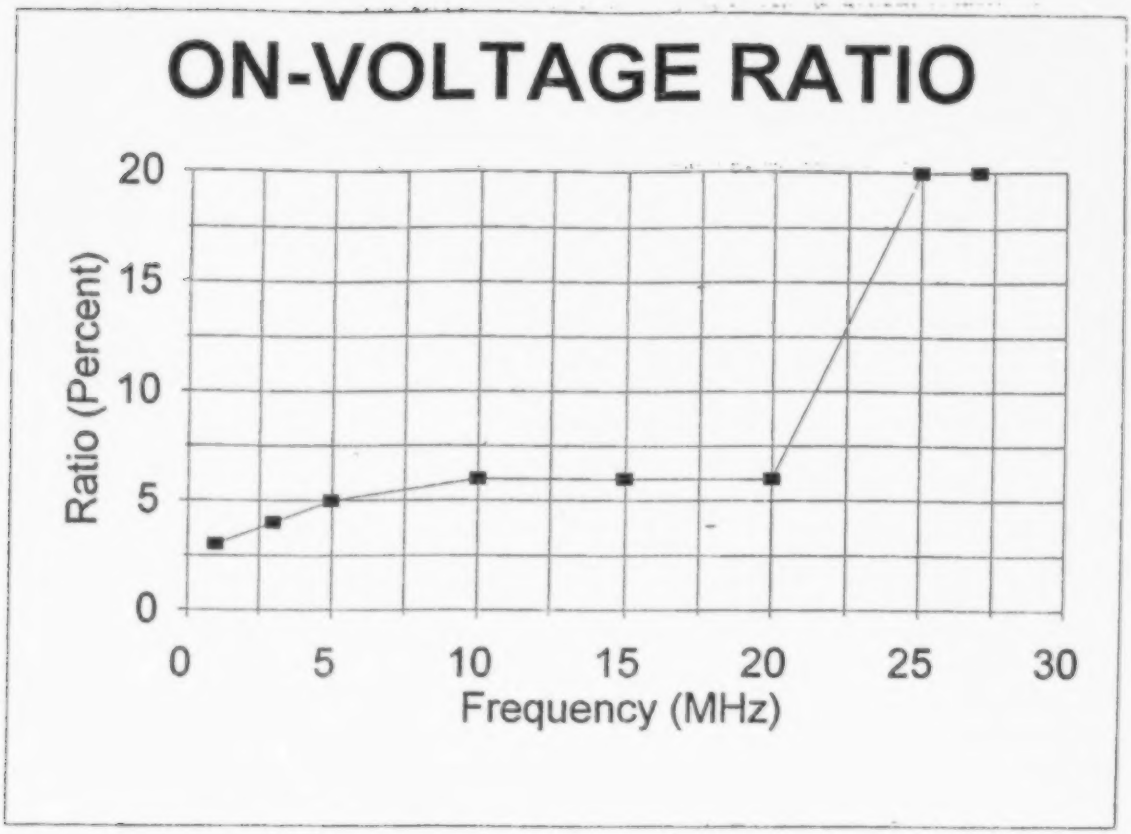
ON-VOLTAGE RATIO
Ratio (Percent)
20
15
10
5
0
0
5
10
15
20
25
30
Frequency (MHz)

Figure 3.

A direct measurement of the capacitance across the antenna terminals was made for the switch open state using a capacitance meter and yielded a value of about 50 pF. This suggests that the data sheets seriously underestimate the output capacitance.

The antenna switch can be placed in the closed position by switching the unit On and placing the jumper shunt across JP-1. For these measurements, the series resistor, R, was set at 75 Ohm. Figure 3 shows the upper limit to the ratio of the voltage across the switch to that from the signal generator as a function of frequency. Stray signals and breakthrough were quite a problem and especially at frequencies above 20 MHz. It appears that the impedance of the switch is probably less than 5 Ohm over the band. The static dc resistance was measured to be 1.6 Ohm. This is much lower than that given in the data sheets and may be associated with the higher output capacitance; it suggests a FET channel with a larger area than normal.

Finally, the unit was tested with a radio receiver. It was placed between the antenna, which was approximately a quarter wavelength at 20 MHz, and the receiver antenna terminal (50 or 400 Ohm). It turned out that the transitions of the square wave were sufficiently sharp to produce signals that could be heard directly at the receiver output. However, when the oscillator was set to low rates, the unit heavily modulated the background noise. When strong signals were present, the AGC tended to suppress the modulation.

Temperature testing consisted of applying a heat gun to the oscillator and MOSFET, which raised the chip temperature to over 60 C: no changes in frequency were apparent. Nor were any significant changes noted in the switching action. Low temperature operation was tested by leaving the unit in the freezing compartment of a refrigerator for two hours before measuring the oscillator frequencies. Some cursory testing of the open and closed circuit switch impedances was also made before the unit warmed up and no significant changes were noted.

APPENDIX A

A SWITCHED RADAR CALIBRATION ANTENNA

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1. INTRODUCTION

The antenna under consideration comprises a stationary half wave dipole with a switch at its centre. When the switch is closed, the antenna acts as a half wave dipole and its scattering cross section is large. When it is open, the antenna scatters as if it were two quarter wavelength rods and the scattering cross section is small. Thus periodic switching will amplitude modulate the Radar Cross Section (RCS). The effect on the radar signal is to create sidebands about the main return and these can be made to lie outside the clutter band. Thus the return from the antenna appears similar to that from moving targets. The apparent Doppler of these targets can be controlled by varying the switching frequency.

The main interest in the antenna is for the calibration of HF radar with vertically polarized signals but, since switching can also be accomplished at microwave frequencies, the concept is also applicable to microwave frequencies. (In this case, the dipole might be replaced by a switchable horn antenna.) Because the cables for carrying the control signals tend to affect the antenna pattern, an optically controlled system seems to be preferable. It may be possible to implement the switch using only optical power with a photo-voltaic detector and a FET. To mimic the Doppler at HF from a target moving at 100 m/s, switching frequencies of about 20 Hz are required.

There are several aspects to the HF antenna that require some study. Firstly, the scattering of the antenna should be examined to determine the effects of less than ideal switching and especially the coupling between the two quarter wave elements when the switch is nominally open. Secondly the switch itself should be optimized and this needs some analysis and a careful switch design. Thirdly the effect of the modulation itself on the radar needs to be determined to establish the effects of multiple sidebands and the possibility of using a modulation other than a square wave.

These problems have been addressed to some extent in the literature and it is the intent of this report to integrate the present information into one document and to present the details of existing analyses as well as to expand on the topic.

2. Scattering from the Dipole

When the switch is closed, the dipole behaves more or less as a normal half wavelength dipole antenna. The incident field causes currents to flow from one element to the other through the switch and these currents re-radiate a signal back to the radar receiver. A calculation of the RCS in this state requires a determination of the current distribution along the dipole and a calculation of the radiated far field. Fortunately these quantities can be estimated readily from existing theory, such as that provided by Ruck et al, 1970. For example, using a simplified theory in which the current distribution is sinusoidal, it can be shown that, when the scattering is at broadside and the polarization vectors of the transmitting and receiving antennas are parallel to each other and perpendicular to the incident wave-vector, the RCS, σ, is a maximum and is given approximately by

$$\sigma \sim 0.86\lambda^2$$

where λ is the wavelength. However, a more rigorous approach changes the factor of 0.86 to 0.99. The gain of the ideal half wave antenna under these conditions is a maximum of 1.64 or 2.15 dB.

The simple approach assumes a shorted dipole but, if the switch is not ideal when closed, there will be some loss. However, this can be estimated using an equivalent circuit method noting that the radiation resistance of the antenna, R_r, is approximately 72 Ω. Therefore, if the switch is purely resistive with a resistance R_S, the received current through the switch and all points on the antenna will be reduced by a factor of

$$\frac{R_r}{R_r+R_S}$$

and the re-radiated power by the square of this factor. For example, if the switch resistance is 10 Ω, there will be a reduction in the RCS by about 1.3 dB.

When the switch is open, the antennas are nominally disconnected and scatter weakly as short dipoles. However the theory is not as complete. In the case of an ideal switch, the equivalent circuit approach would suggest that no current flows and that the RCS of the antenna is zero. Therefore it is necessary to estimate the weak scattering from each element when isolated. Ruck et al, 1970, show that the maximum RCS of a short length of conducting wire is given by:

$$\sigma \sim \frac{4\pi}{9}\frac{k^4h^6}{(\ln(4h/a)-1)^2}$$

where h and a are the half length and the radius of the wire and k, the angular wavenumber, is equal to $2\pi/\lambda$. For a quarter wave antenna, this becomes:

$$\sigma \sim \frac{\pi^5\lambda^2}{36864(\ln(\lambda/(2a))-1)^2} = 0.0083\frac{\lambda^2}{(\ln(\lambda/(2a))-1)^2}$$

Therefore the RCS of two practical isolated quarter wave elements is much less than one percent of the RCS of a half wave dipole.

This result suggests that, for practical purposes, an equivalent circuit for the half wave dipole can be employed even when the switch is open. For example, if in the open position the switch were entirely resistive with a resistance of 500 Ω, the current through the switch would be reduced by a factor of 0.126 over that of a shorted half wave dipole. The RCS is then reduced by the square of this, namely 0.016 or 18 dB: this is still much greater than the return from the isolated quarter wave elements.

3. The Switch

Various types of switch may be used depending on the application. For high powers back to back PIN diodes may be preferred, especially those activated optically as described by Albares et al. 1999 or Sun et al, 199?. However, the device known as the photo-injection pin (PIPIN) switch requires a significant optical power for biasing the diodes. At low power, the optically switched FET appears to be the device of choice (Albares et al, 1999, Sun et al, 1996, Sun et al, 1997). Here the photo detector can be separate or integrated within the transistor structure: such an integrated scheme has been discussed by Freeman et al, 1992. These special devices do not seem to be available commercially but can be made from discrete components as has been demonstrated by Sun et al, 1996.

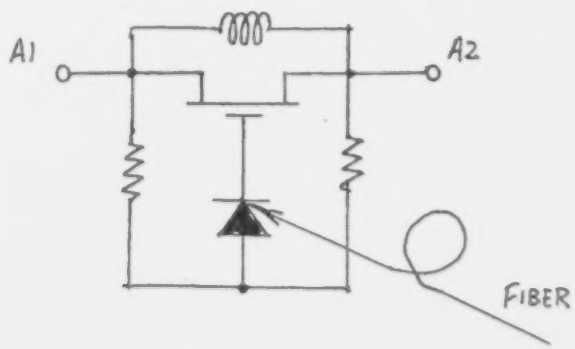

Figure 1
FET Switch

The circuit of a photo-voltaic FET switch for antenna applications is as shown in Figure 1. The resistors from the source and drain to the gate can be large compared to the antenna radiation resistance of a half wave dipole and can be neglected in the analysis of the antenna RCS. The inductor is designed to resonate with the capacitance of the FET and other stray capacitance while the switch is open. Therefore, to a first approximation, the switch always appears to be resistive.

The performance of the switch is summarized by its insertion loss (IL) and its isolation. The insertion loss is defined by the loss of power that occurs in a matched load due to the switch. When the switch is closed and introduces a resistance, R_0, the power in the load is:

$$P = \frac{e^2}{(2Z_0+R_0)^2}Z_0$$

where e is the rms emf of the generator and Z_0 is the impedance (assumed resistive) of the generator and its matched load. This must be compared with the power when the switch is absent, i.e. when $R_0=0$. The ratio of these powers defines the insertion loss and can be expressed in decibels as

$$IL = 20 \log_{10}\left(1+\frac{R_0}{2Z_0}\right)$$

This is the result of Sun et al, 1996.

The isolation provided by the switch is calculated again for the case of matched generator and load but with the switch open. Stray capacitance is likely to reduce the isolation unless it is canceled by inductive reactance. Sun et al, 1996, employ a model in which the switch is represented by a pure capacitor, C, and a resistor, R_0, in series while the inductor is represented by a pure inductor, L, and its resistance, R_L. It is easily shown that, when this tuned circuit is resonant, its dynamic resistance, R_D, is given by:

$$R_D = \frac{L}{C(R_0+R_L)}$$

The isolation is again defined by the ratio of the power dissipated in the load when the switch is absent to that when it is present in the open position. This ratio can be expressed in decibels as

$$Isolation = 20 \log_{10}\left(\frac{2Z_0+R_D}{2Z_0}\right)$$

Note that at resonance the inductive and capacitive reactances are equal so that with frequency, f, we have

$$f^2 = \frac{1}{4\pi^2 LC}$$

Combining the previous three equations to eliminate R_D and L, we have (in decibels)

$$Isolation = 20 \log_{10}\left(1+\frac{R_0 f_c^2}{2Z_0 f^2}\right)$$

where f_c is defined by

$$f_c^2 = \frac{1}{4\pi^2 C^2 R_0 (R_0 + R_L)}$$

This result is slightly different from that of Sun et al, 1996, because it is felt that the channel resistance and another series resistance introduced by these authors should not be separated. Evidently f_c can be regarded as a figure of merit for the switch.

Sun et al, 1996, employed a commercial GaAs photo-voltaic cell and a JFET, apparently of a special type. They achieved a resistance of a few Ohms for the closed position and greater than 30 MΩ in the open position. The capacitance in the open position was about 0.3 pF and the switch was operated at 790 MHz to yield an isolation of 57 dB. The rise time of the switch was about 20 μs and the fall time about 2 μs.

4. Sidebands

The radar receiver is sensitive to the spectrum of the received signal. The modulation of the RCS by a square wave causes a line spectrum corresponding to its amplitude modulation. It is useful to calculate the apparent RCS associated with each of the sidebands as well as that from the carrier. The model for this is a periodic sinusoidal signal with an amplitude that also varies periodically but much more slowly in the form of a square wave from a high value of A_1 to a low value of A_2. These amplitudes are proportional to the square roots of the high and low values of the RCS.

The spectrum of the signal can be deduced using the Fourier series method. It is simplified if one period of the modulating square wave signal contains an integer number of periods of the carrier signal. Spiegel, 1970, presents the Fourier series and this can be simplified. Thus the signal,

f(θ), can be represented as

$$f(\theta) = \frac{a_0}{2} + \sum_{n=1}^{\infty} (a_n \cos(n\theta) + b_n \sin(n\theta))$$

The coefficients a_n and b_n are found from:

$$a_n = \frac{1}{\pi} \int_0^{2\pi} f(\theta) \cos(n\theta) d\theta$$

$$b_n = \frac{1}{\pi} \int_0^{2\pi} f(\theta) \sin(n\theta) d\theta$$

The power can be expressed as the sum of the power in each side-band using Parseval's theorem:

$$P = \frac{a_0^2}{2} + \sum_{n=1}^{\infty} (a_n^2 + b_n^2)$$

The mean value of the signal is clearly zero so that $a_0=0$. If the carrier signal frequency corresponds to index n=m, we can find the coefficients a_m and b_m corresponding to the carrier in the spectrum. Thus

$$a_m = \frac{1}{\pi} \int_0^{\pi} A_1 \sin(m\theta) \cos(m\theta) d\theta + \frac{1}{\pi} \int_{\pi}^{2\pi} A_2 \sin(m\theta) \cos(m\theta) d\theta$$

$$b_m = \frac{1}{\pi} \int_0^{\pi} A_1 \sin(m\theta) \sin(m\theta) d\theta + \frac{1}{\pi} \int_{\pi}^{2\pi} A_2 \sin(m\theta) \sin(m\theta) d\theta$$

An evaluation of these integrals yields:

$$a_m = 0$$

$$b_m = \frac{A_1 + A_2}{2}$$

The first side-bands correspond to n=m-1 and n=m+1. For a sideband separated by s from the carrier we have:

$$a_{m+s} = \frac{1}{\pi}\int_0^{\pi} A_1 \sin(m\theta)\cos((m+s)\theta)d\theta + \frac{1}{\pi}\int_{\pi}^{2\pi} A_2 \sin(m\theta)\cos((m+s)\theta)d\theta$$

$$b_{m+s} = \frac{1}{\pi}\int_0^{\pi} A_1 \sin(m\theta)\sin((m+s)\theta)d\theta + \frac{1}{\pi}\int_{\pi}^{2\pi} A_2 \sin(m\theta)\sin((m+s)\theta)d\theta$$

These can be integrated using elementary trigonometric identities, such as

$$2\sin(m\theta)\cos((m+s)\theta) = \sin((2m+s)\theta) + \sin(s\theta)$$

Thus we have:

$$a_{m+s} = \frac{A_1 - A_2}{\pi}\left(\frac{1}{2m+s} + \frac{1}{s}\right) \quad \textit{for } s \textit{ odd, } 0 \textit{ otherwise}$$

$$b_{m+s} = 0$$

When s is odd, the two terms in a_{m+s} correspond to the contributions from positive and negative frequency components in an analysis based on a complex transform.

Since m>>s in cases of practical interest, the power in the s-th (odd) side-band, P_s, is given by

$$P_s = \left(\frac{A_1 - A_2}{s\pi}\right)^2$$

In terms of the corresponding values of the RCS, σ_1 and σ_2, the RCS of the carrier, σ_0, is given by

$$\sigma_0 = \left(\frac{\sqrt{\sigma_1} + \sqrt{\sigma_2}}{2}\right)^2$$

Clearly when $\sigma_1 = \sigma_2$, there is no modulation and the result is to be expected. The RCS of the side-bands for s odd is given by

$$\sigma_s = \left(\frac{\sqrt{\sigma_1} - \sqrt{\sigma_2}}{s\pi}\right)^2$$

This can be cast into a form in which the RCS is expressed in dBsm (indicated by primes), i.e.

$$\sigma_s' = 20 \log\left(\frac{\sqrt{\sigma_1}(1-\sqrt{\sigma_2/\sigma_1})}{|s|\pi} \right) = \sigma_1' + 20 \log\left(\frac{1-10^{-R/20}}{|s|\pi} \right)$$

where $R=\sigma_1'-\sigma_2'$ and s is odd.

These expressions for the side-band RCS clearly indicate that the smaller of the two RCS values is not important as long as it is reasonably small.

5. Conclusion

It appears very likely that a switchable calibration antenna could be implemented successfully at HF using commercial components and including a fibre-optic link for switching. The fibre-optic part of the device is standard. It might also be worthwhile studying the same type of system at X-band using a horn and placing the switch at an appropriate position in the waveguide to cancel stray capacitance.

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APPENDIX B

FIBER OPTICS

Signals from a LED or a laser can be carried down an optical fiber where they can be converted into electrical signals by a detector. The fiber itself introduces a only a small loss because it relies on the principle of internal reflection and the material, which is usually glass and sometimes plastic, is very pure so that scattering is minimal. If the interface responsible for the total internal reflection were outside of the fiber and were air, the inevitable surface imperfections would create serious losses so that the fiber typically consists of a central core surrounded by a cladding of material of lesser refractive index which is bonded to the core. Outside the cladding there is some plastic buffer material for protection and additional mechanical strength. Attenuations of the order of a few dB/km are typical in glass fibers but are much higher for plastic fibers.

The fiber can be designed for multimode or single mode transmission. In the multimode case, there are a variety of what amount to discrete waveguide modes. These create dispersion in the signal and this limits the bandwidth. However, the fiber can transmit signals arriving over a wide range of angles to the axis (numerical aperture) which makes matching of the fiber to signal sources and receivers easier. Multimode cable is typically thicker than single mode cable.

Single mode cable has a core with a diameter only a few times the wavelength and does not suffer from the dispersion arising from multiple modes. The bandwidth is much higher in the gigahertz range and is not required for this application. It will not be discussed further.

To limit the dispersion in multimode cable, a graded refractive index can be used so that the paths of extreme rays tend to pass through material of lesser refractive index and the group velocity of these waves is higher. This helps to compensate for the dispersion. In contrast, in the step index cable all rays are reflected at a discrete interface. The dispersion in graded index multimode cables is typically of the order of a few ns/km. The bandwidth is typically 0.5 GHz-km.

When a fiber is used for communications over long distances attenuation is important. It can occur through absorption and scattering from impurities in the material of the core or in the cladding near the reflecting interface. Thus there are absorption bands that must be avoided, such as that due to OH^+ ions. The scattering is primarily of the Rayleigh type and falls off as the fourth power of the wavelength. Outside the absorption bands Rayleigh scattering predominates and therefore long wavelengths are to be preferred. However, the length of fiber in the current application is only of the order of 100 m so that any of the usual wavelengths will be appropriate.

The wavelengths used for communications are 850 nm, 1300 nm and 1550 nm and devices may be purchased for these wavelengths. It should be noted that plastic fibers, which should be avoided because of their high attenuation of the order of 1 dB/m, are generally used with wavelengths of 665 nm. Because the attenuation in multimode glass fibers should not be an issue,

any of the wavelengths could be employed depending on the availability of devices. The cable itself is not specific to the wavelengths noted.

There are various glass fiber multimode cables. These are the 50/125, the 62.5/125, the 85/125 and the 100/140. The numbers refer to the diameter of the core and the cladding respectively in micrometres. The most popular cable is the 62.5/125.

APPENDIX C

ANTENNA SWITCH DESIGN

The antenna switch could take various forms, such as a Micro Electro-Mechanical Switch (MEMS), more conventional mechanical switch (e.g. a reed relay) or a semiconductor switch. Whatever the form of the switch, it should have a low impedance in the on condition and a high impedance in the off condition. Furthermore, it should not consume much power when it is activated and should be capable of many switching cycles without degradation as well as being capable of switching at speeds as high as 20 Hz.

When a switch is in the on state, its impedance at HF is typically low and is almost entirely resistive. Stray capacitive reactance in parallel with this resistance can be made negligible by keeping conducting traces short. Similarly the series inductive reactance of the conductors can be made small by keeping the wiring thin and the lengths of the wires short.

In the off state, the impedance is principally a capacitive reactance corresponding to the gap across the contacts of a relay or their equivalent. At sufficiently high frequencies, there may also be a significant resistive component. Series inductance will usually be cancelled by the capacitive reactance. It is a simple matter to estimate various capacitive reactances and inductances as is demonstrated in Appendix D. For a mechanical relay, the series inductance will be less than 1 μH while the capacitance in the off state should not be more than a few pF. This suggests that switching at GHz frequencies is possible.

Mechanical relays typically exhibit a very low resistance often much less than 1 Ω in the on position and their impedance in the off position is controlled by the capacitance across the contacts and between the conductors connected to them. However, they require upwards of 100 mW to hold the switch in one of its states. While not eliminating the mechanical switch as a suitable candidate (especially a MEMS), this makes the semiconductor switch attractive because it can be operated with extremely small powers. However, the output capacitance of the semiconductor switch may be comparatively large. The present design utilises a semiconductor switch because we have not been able to obtain MEMS or determine their open-circuit capacitance.

The choice of a semiconductor switch is constrained by its resistance when it is in the on state and by the impedance when it is off. These constraints are not independent because the area of the channel of a FET controls both. A study of the Motorola data sheets indicates that the BS107A, which is a MOSFET, may be a suitable choice. This is stated to have an on-resistance of about 5 Ω and an output capacitance as low as 15 pF. There are a number of other switching FET's but most of these can be eliminated because either the output capacitance in the off state is too high or the resistance in the on state is too low. The BS107A is a N-channel enhancement device. This implies that it is in the off state when the gate and source are at the same potential

and that it can be turned on by applying a positive voltage to the gate. Thus it is compatible with TTL signal levels and specifically the 555 oscillator. In comparison, JFET N-channel depletion devices would require a negative voltage to turn them off and this would necessitate two batteries.

The output capacitance of the semiconductor switch tends to limit the high frequency performance so that this type of switch can only be used for broadband antenna switching at HF. For example the reactance of a 20 pF capacitance at 25 MHz is about 320 Ω. Thus it is of the same order as the radiation resistance of a half wave dipole. This capacitive reactance can be cancelled by placing an inductor across the switch but cancellation is only effective over a finite bandwidth.

APPENDIX D

ESTIMATION OF CAPACITANCE AND INDUCTANCE

First we can estimate the capacitance of an isolated thin wire of length, l, and radius R. The electric field about the wire will be symmetric about the wire axis and, near its centre will diverge radially. Thus, near the wire and in its centre region, the field behaves as if it arises from an infinite cylinder. Thus it can be estimated accurately. As the radial distance from the wire increases, the effect of the finite length of the wire starts be become important and, at a distance approximately equal to its length, the form of the field changes from an inverse radius dependence to an inverse square dependence, as if it were to arise from a point change or small sphere.

The capacitance of an object can be interpreted in terms of the energy in the electric field. This is proportional to the volume integral of the square of the electric field. If the field diverges as $1/r^2$, as for a point change, it is clear that, since the volume elements contain a factor of r^2, the field in the neighbourhood of the object makes a far more important contribution to the capacitance than the field far away. Thus, when estimating the capacitance, the distant field can be neglected entirely. The results obtained in this way can be justified if they are very insensitive to the cut-off point in the calculation.

Applying the method to the thin wire, we can arbitrarily decide that the field is similar to that for a cylinder out to a radius equal to the length of the wire. The result is then that its capacitance, C, is given by:

$$C = \frac{2\pi\varepsilon\varepsilon_0 l}{\ln(l/R)}$$

where ε and ε_0 are the relative permittivity of the medium embedding the wire and the permittivity of free space respectively. Because of the logarithmic dependence of the cut-off point in the integration to find the electric potential, the constraints within the estimation method are satisfied.

Similarly the method can be applied to an estimation of the inductance of a length of wire but here a volume integral over the square of the magnetic field is required. The result is that its inductance, L, is given approximately by:

$$L = \frac{\mu_0 l \ln(l/R)}{4\pi}$$

where μ_0 is the permeability of free space. For typical thin wires or circuit board traces in vacuum or air we find that for a 1m length of wire, the capacitance is about 10 pF and the inductance is

about 1 μH.

The capacitances and inductances estimated above apply to a situation for example where a wire is separated from another conductor by a distance of the order of its length. If a ground plane is much closer than this but is still a large distance compared to the wire radius, the field will be increased by the image of the wire in the plane and the integral can be performed over a shorter distance. Thus, for the thin wire parallel to a ground plane at distance r', we have:

$$C = \frac{4\pi\varepsilon\varepsilon_0 l}{\ln(r'/r)}$$

It is worth noting that these results can also be applied to a half wave dipole (each element is a quarter wavelength in length). In this case the capacitance of the dipole is about one half of that for each element while the overall inductance is twice that of each element. The resonant frequency of the system can be estimated as if it were a tuned circuit and indeed, by noting that the velocity of light,c, is related to the permittivity and permeability of free space by:

$$c^2 = \frac{1}{\mu_0\varepsilon_0}$$

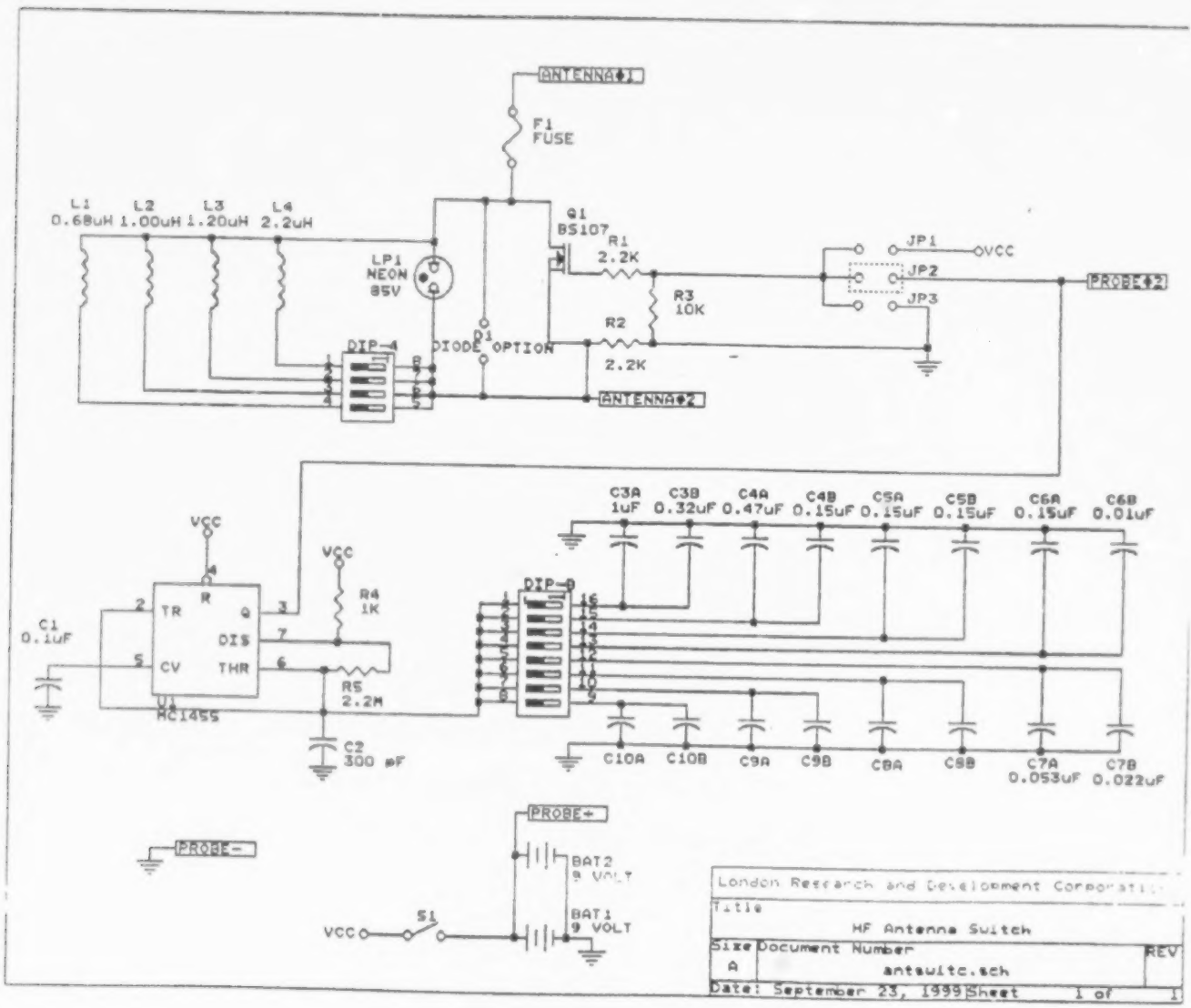
we find that

$$f^2 = \frac{32}{4\pi^2}\frac{c^2}{\lambda^2}$$

Thus, apart from a numerical factor of the order of unity, the correct result is obtained.

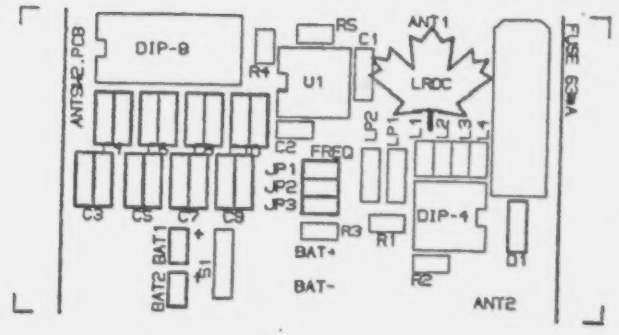
APPENDIX E

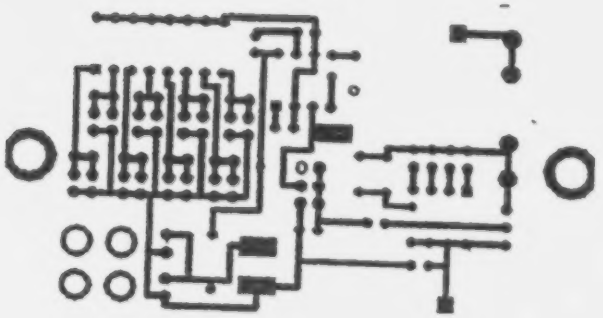
CIRCUIT DIAGRAM

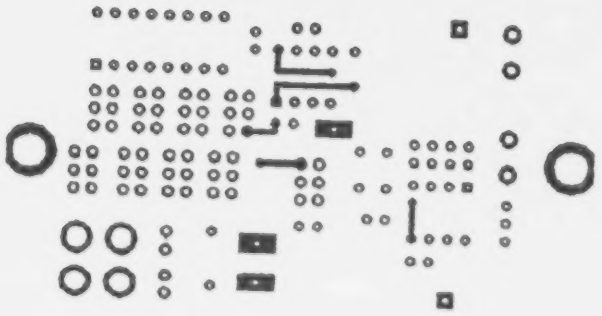


APPENDIX F

PRINTED CIRCUIT LAYOUT

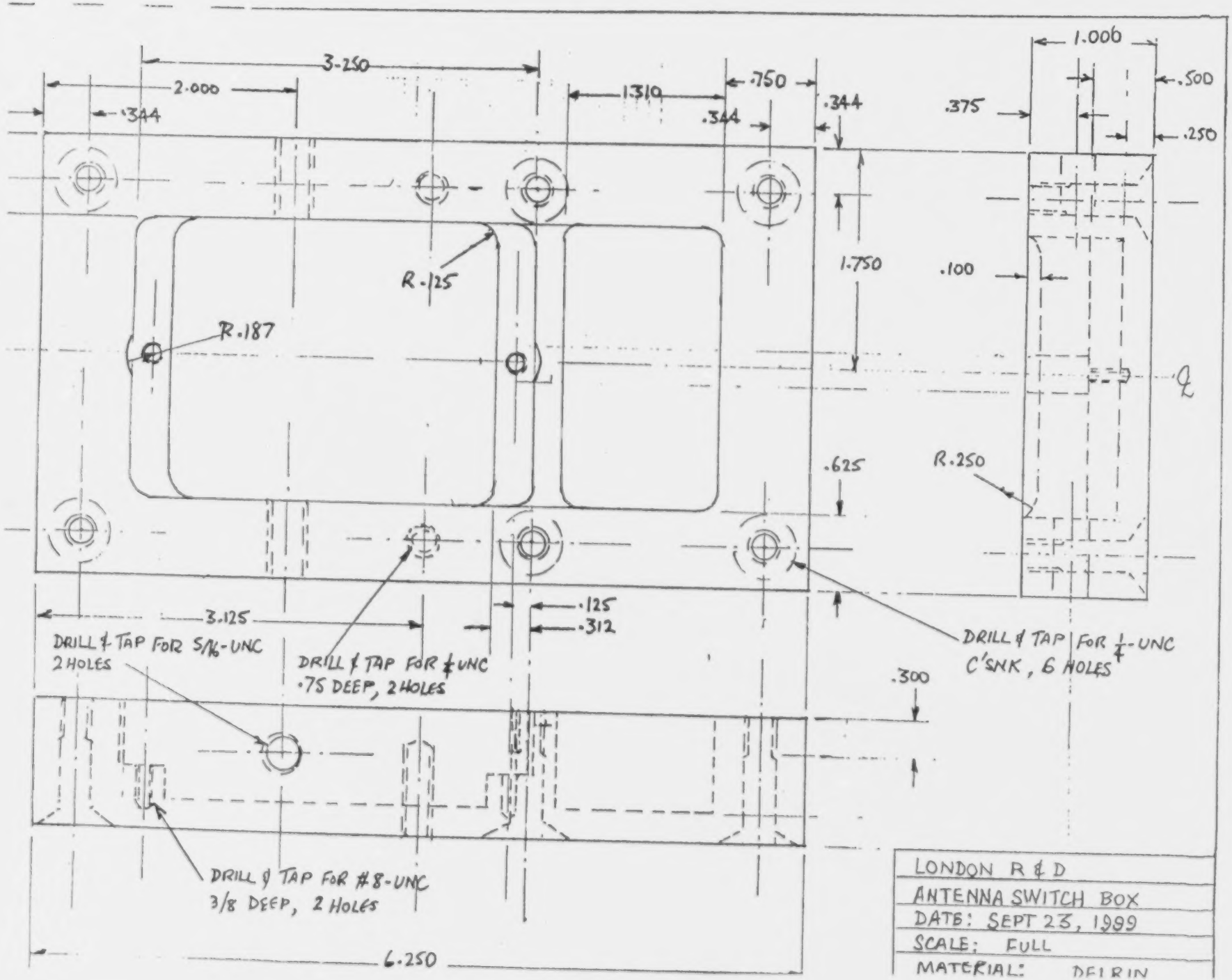






APPENDIX G

MECHANICAL DRAWINGS


3.250
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1.750
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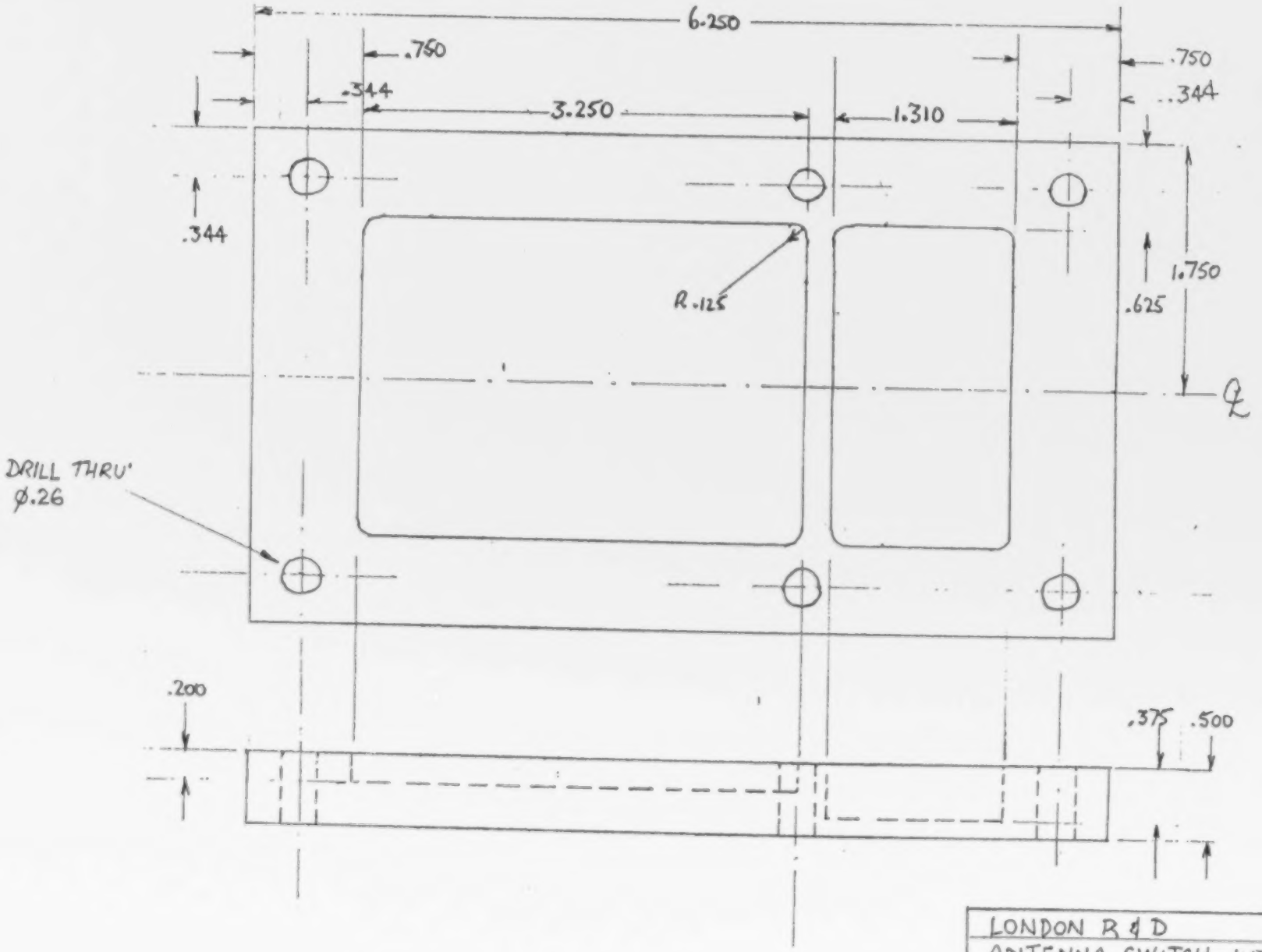
R.187
R.250
.625
3.125
.125
.312
.300
DRILL & TAP FOR 5/16-UNC
2 HOLES
DRILL & TAP FOR 1/4 UNC
.75 DEEP, 2 HOLES
DRILL & TAP FOR 1/4-UNC
C'SNK, 6 HOLES
DRILL & TAP FOR #8-UNC
3/8 DEEP, 2 HOLES
6.250
LONDON R & D
ANTENNA SWITCH BOX
DATE: SEPT 23, 1999
SCALE: FULL
MATERIAL: DELRIN



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DATE: SEPT 23, 1999
SCALE: FULL
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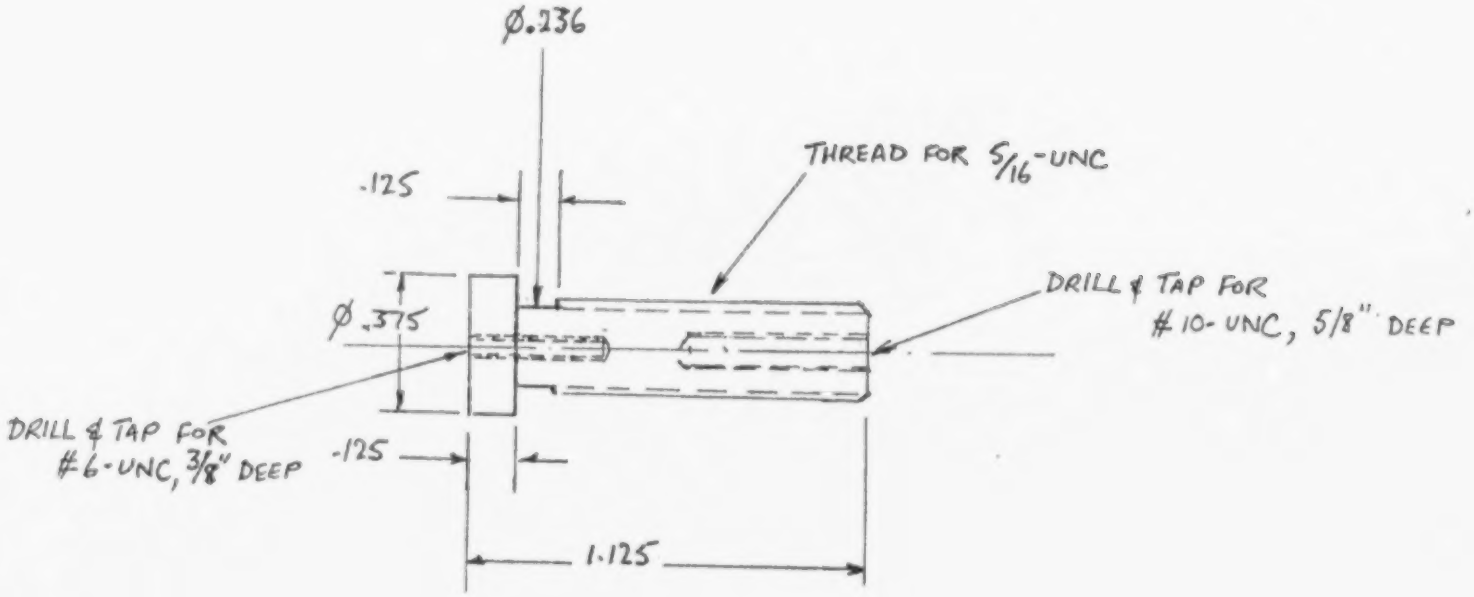
QUANTITY: 2 PER UNIT



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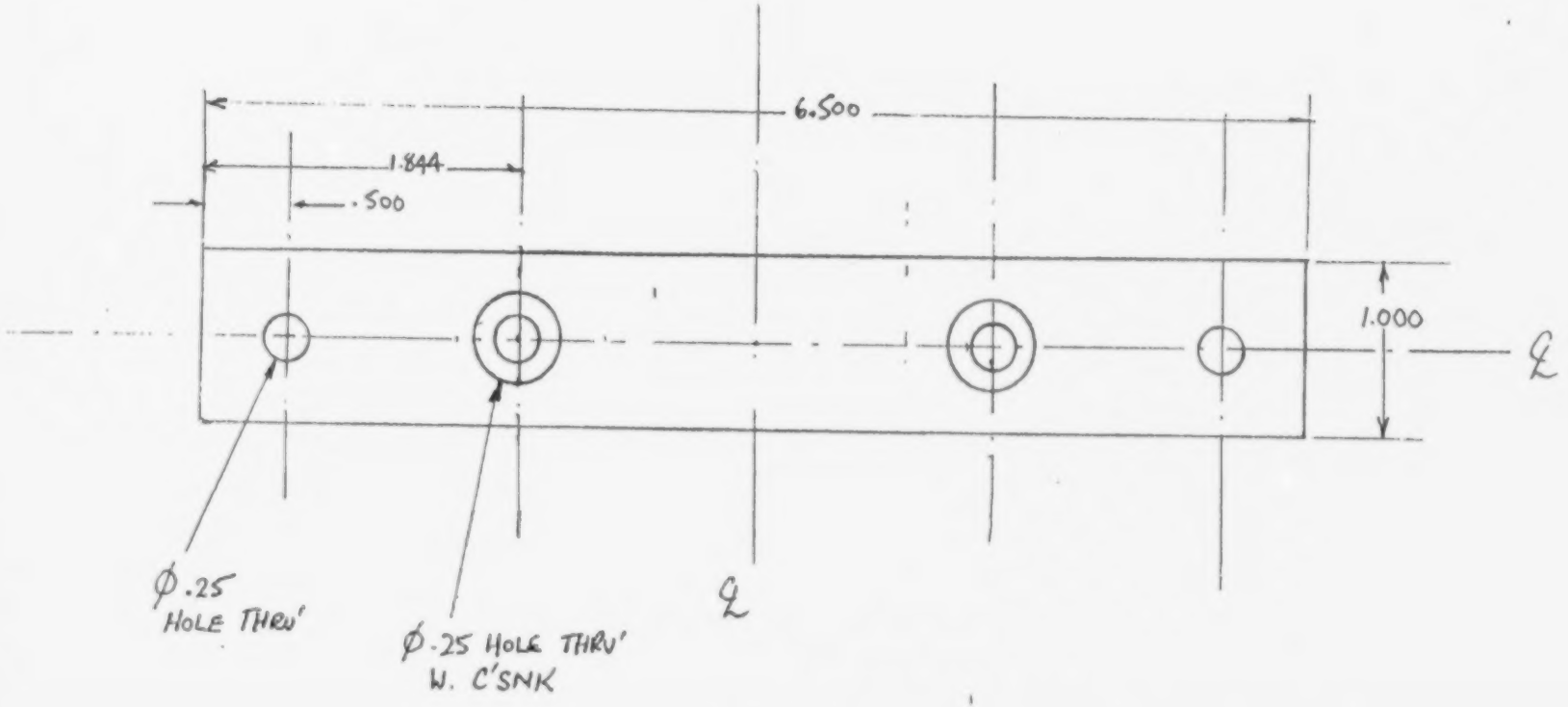
DATE: SEPT 23, 1999
SCALE: 2:1

NOTE: STANDARD DELRIN 1/4" STOCK



LONDON R & D
ANTENNA SWITCH MOUNT
DATE: SEPT 23, 1999
SCALE: FULL
MATERIAL: DELRIN

APPENDIX H

SUMMARY OF DATA SHEETS

DATA FOR 2N7000	
Type	TMOS FET Transistor
Maximum Drain-Source Voltage	60 V
Maximum Drain-Gate Voltage	60 V
Maximum Drain Current	200 mA (Continuous)
Drain Current for Zero Gate Voltage	1 μA
Drain-Source On-Voltage	2.5 V
Static Drain-Source On-Resistance*	5 Ω
Output Capacitance**	25 pF
Reverse Transfer Capacitance	5 pF

* Measurements of the static on-resistance tend to be significantly less than 5 Ohm.

** Measurements of the output capacitance suggest that the output capacitance is larger than that stated in the Motorola data.

DATA FOR BS107	
Type	TMOS Switching Transistor
Maximum Drain-Source Voltage	200 V
Maximum Source-Gate Voltage	20 V
Maximum Drain Current	250 mA (Continuous)
Drain Current for Zero Gate Voltage	0.03 μA
Drain-Source On-Voltage	1 to 3 V
Static Drain-Source On-Resistance	4 to 20 Ω
Output Capacitance	20 pF
Reverse Transfer Capacitance	3 pF

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Tunaley, J.K.E.

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In the setup and testing of High-Frequency (HF) radars, it is often necessary to have a target of known cross section. Separating the target reflections from low-Doppler-shift clutter is usually difficult, so a useful target will have an oscillating cross section. This is most easily done by placing a switch between components of the target so the target cross section is high when the switch is closed and low when it is open. This report describes the construction of an all-solid-state switch which performs the switching function for radio frequencies in the range 1-27 MHz. The switch is based on a 2N7000 MOSFET transistor, and can run for about 10 hours on 2 9-V transistor batteries. Across the 1-27 MHz range, its open-state impedance is >470 Ohms while the closed-state impedance is <5 Ohms. The report also discusses alternate approaches to design of switches HF radar targets, and also makes calculations of target cross sections when the switch is placed at the centre of a resonant dipole.

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High-Frequency radar
Radar targets
HF switch

#512191